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INPUT CURRENT SHAPED AC-TO-DC CONVERTERS

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1. INTRODUCTION

In switched-mode power converters, such as ac-to-dc regulators, attention is usually focused on their voltage-to-voltage conversion properties. That is, an input voltage and an output voltage are specified, and the output current is then determined by the load. The resulting input current remains unspecified, except insofar as it affects the desired efficiency.

Increasing attention to the waveshape of the input current is now required. The general objective of the present work is to investigate input current shaping techniques for ac-to-dc converters, with particular attention paid to input (power line) frequencies much higher than normal, up to 20 kHz. At 20 kHz, some of the simpler methods may be stronger contenders than usual.

1.1 Necessity for Input Current Shaping

There are two major reasons why increased attention needs to be given to input-current shaping for ac-to-dc power converters, both of which lead to the conclusion that the input current waveshape should be the same as that of the input voltage, and in phase with it. Since an ac input voltage is usually sinusoidal, this means that the current should also be sinusoidal and in phase with the voltage. The conclusion is also true for dc input voltage: the input current should also be dc. For sinusoidal input voltage, the degree of nonideality of the current waveform can be expressed by the power factor PF, defined as

$$PF = \frac{I_F}{I_T} \cos \phi$$

where I_F is the rms value of the fundamental component of the input current, I_T is the rms value of the total input current, and ϕ is the phase angle between the fundamental component of the input current and the input voltage. It is desirable to have a PF as close to unity as possible.

The first reason why the input current should be of the same waveform as, and in phase with, the input voltage is because of the requirement for high efficiency. Any current or voltage repetitive waveform may be represented by a Fourier series: only current components of the same frequencies as the voltage components deliver power from source to load; however, all current components cause power loss in parasitic resistance. Therefore, any current component of a frequency not present in the voltage waveform contributes to power loss in parasitic resistance but does not contribute to power delivered to the load. Parasitic loss resistance is present in the load (converter input), in the line connecting source to load, and in the source itself (source resistance). For a given current waveform (given parasitic loss), maximum power is delivered from source to load when the current waveform is in phase with the voltage waveform (all current components in phase with the corresponding voltage waveforms). Parasitic power loss in excess of the minimum leads not only to loss of electrical efficiency, but also to poor utilization of source and line capacity.

The second reason why the input current waveform should be the same as the voltage waveform is that otherwise unnecessary noise is generated, both radiated and conducted.

In a converter operating from a dc input voltage, any ac component of input current can be considered as noise. There are two sources of such noise: reflected load current, and internally generated ac currents caused by the switching mode of operation of the converter, which consist of the switching frequency and its harmonics. In the presence of ac load currents, there may also be sideband frequencies formed with the switching frequency and its harmonics.

In a converter operating from an ac input voltage, there are input noise currents not only from the load and from the internal switching action, but also harmonics of the line frequency caused by rectification of the input voltage. These harmonics can be large, as in the common half-wave, full-wave, or bridge rectifiers followed by a large reservoir capacitor.

There are several ways to eliminate or alleviate the problems of undesirable input current components, as discussed in the next section.

1.2 Solutions to the Problem

Possible solutions to the problem of undesirable input current waveforms range from primitive to sophisticated.

In a dc input converter, ac load and switching current components in the input are filtered by a capacitor. The simplest brute-force solution is to make the capacitor large enough to reduce the residual ac input current components to an acceptable level. A more elegant solution is the use of a "supercapacitor," in which a relatively small capacitance is effectively multiplied by an active circuit to produce a much larger value. The ac current in the capacitor is sensed and multiplied by a switched-mode bridge circuit, and added in parallel with

the load so that the total filtered ac shunt current is much larger than it would be for the capacitor alone [1].

There is a wide variety of passive and active filtering techniques for use on both dc and ac inputs. Examples are low-pass filters, harmonic traps, harmonic compensation or injection, and shunt filters, all of which tend to be bulky and heavy. Ferroresonant transformers are larger than a conventional transformer of the same rating, and also require a fairly large resonating capacitor. Multiphase transformers can give cancellation up to high harmonic orders, but also must be considerably larger than a conventional transformer owing to poor utilization of their multiple secondaries.

All the above approaches are brute-force in the sense that they reduce noise currents already present at the converter input. More elegant techniques attempt to reduce such noise input currents in the first place.

The simplest such method for reducing rectification input current noise is to employ an inductor between the rectifiers and the reservoir capacitor. This is the conventional inductor-input filter, in which the rectifier conduction angle is increased so that continuous conduction occurs and the input current is sinusoidal.

Another method shapes the ac line input current to be of the same waveform as the line voltage by a pulse-width-modulation regulation technique [2,3,4]. Thus, line harmonic currents are prevented from occurring at the input rectifiers, and the remaining switching frequency noise is much easier to suppress. Also, the line current regulation can be designed to control current surges. This method effectively can incorporate line current waveform regulation and output voltage

regulation in the same pwm feedback loop, by controlling both frequency and duty cycle regulation of the power switch. However, the control must be constant over a half-period of the line frequency in order to preclude the generation of line frequency harmonics, which greatly degrades the response to load and line transients because of the required low bandwidth of the control loop. As a result, a two-stage switched-mode converter is often used, with one shaping the line input current waveform and the other providing wideband line and load regulation from the low-bandwidth dc output of the first converter.

A more sophisticated approach is to operate a dc-to-ac inverter in the reverse direction, as an ac-to-dc "shaped rectifier." A wide variety of switched-mode dc-to-ac inverter topologies exists; in principle any of them can be operated in the reverse direction if the switches are made capable of operating in the required quadrants. Actually, the implementation is considerably more difficult since the ac port, instead of the dc port, is terminated by a low impedance, and the current waveshaping must be forcibly imposed by the inverter rather than being an automatic consequence of a (relatively) high-impedance load.

Conceptually the most general technique is that which implements a generalized transformer, in which input and output can be of different frequencies, including dc. One such scheme [5,6] achieves this in a three-phase system (for which the instantaneous power throughput is constant) with use of only switches. There are no harmonics below the switching frequency, and no energy storage elements except in the filters that eliminate the switching frequency harmonics. Furthermore, the input power factor can be independently controlled.

2. METHODS OF INPUT CURRENT SHAPING

In this section a number of methods of achieving input-current waveshaping are reviewed, some of which are amenable to either single-phase or three-phase inputs.

Two of the most promising solutions, at least for a 20 kHz input, are discussed first: the simple LC filter following the rectifiers, and the resistor emulation method in which the input current waveform is actively shaped to be sinusoidal by appropriate control of a pulse-width-modulated (pwm) regulator. Then, several other methods, including other forms of pwm, active filter, topology modification high-frequency switching, and multiconverter techniques are reviewed as to their principle of operation.

2.1 LC Filter

The simplest form of ac-to-dc converter is a diode rectifier followed by a filter. Usually the rectifier is full-wave, and the filter consists of at least a reservoir capacitor. In many applications a dc-to-dc regulator is also present, and a typical block diagram of such a system is shown in Fig. 2.1.

Although in this case the regulator is independent of the rectifier and filter, to the extent that the filter capacitor can be considered infinitely large, the regulator is included in the present discussion. Since it will normally be present a more realistic comparison can thus be made with the resistor emulation method of input-current waveshaping, in which the LC filter is effectively incorporated into the regulator.

In the present case, however, the regulator can be ignored functionally except insofar as it modifies the loading on the filter.

The rectifiers present a rectified sine wave $e_{r.c.}$ to the filter, and an equivalent circuit of the rectifier and filter is shown in Fig. 2.2. The diode is present because the rectified current $i_{r.c.}$ can only flow into the inductor L . Capacitor C is normally sufficiently large that the filter output voltage v_0 is essentially dc, and only the dc component $i_{a.v}$ of the load current i_0 flows through the inductor.

The desired result is that the power supply input current $i_{i.n}$ should have the same shape as, and be in phase with, the line sinusoidal input voltage $e_{i.n}$. Translated to the equivalent circuit of Fig. 2.2, this means that $i_{r.c.}$ should be a half sine wave in phase with $e_{r.c.}$, over each half cycle.

There is only one circuit parameter open to choice, namely, the value of the inductance L , since the capacitor is effectively infinitely large. The current waveform $i_{r.c.}$ is shown in Fig. 2.3 for several values of L .

If $L = 0$, the circuit degenerates to the familiar rectifier/capacitor filter, in which $i_{r.c.}$ consists of short, high-peak charging pulses, as shown in the upper current waveform in Fig. 2.3. In this case, the dc output voltage v_0 approaches the peak value of $v_{r.c.}$, but this is not shown in Fig. 2.3 in order to accommodate all cases in one diagram. Clearly, short high-peak current pulses do not constitute a desirable result and the inductor is included specifically to widen and lower these pulses, as shown in the second current waveform in Fig. 2.3. This waveform is closer to the desired half sine wave, but still has intervals of zero value.

As the inductance value is increased, at some critical inductance L_c the zero intervals of the current $i_{r.c.}$ disappear, and the current

waveform thereafter becomes continuous, with an ac component whose peak value is less than its average or dc value i_{av} , as shown in the third current waveform in Fig. 2.3. As the inductance value increases further, the current tends to become more and more constant until ultimately, at $L \rightarrow \infty$, i_{rec} becomes constant at i_{av} (see the fourth current waveforms in Fig. 2.3).

The line input current i_{in} is the "unrectified" i_{rec} , in which the second half-cycle waveforms in Fig. 2.3 are inverted. It is seen that no waveform achieves the desired half sine wave shape over a half cycle, regardless of the inductance value. In particular, an infinite inductance L leads to a square wave input current rather than a sine wave. Nevertheless, some acceptable waveform might be achievable.

A thorough analysis of the equivalent circuit of Fig. 2.2 has been done by Schwarz [7], for both the discontinuous and the continuous inductor current modes of operation. The analyses are done numerically, and the results presented graphically. Schwarz defines two descriptors of the input current waveform, the power factor PF, and the form factor $\rho = i_{rms}/i_{av}$ where i_{rms} and i_{av} are respectively the rms value and average value of the inductor current i_{rec} .

Schwarz's results are shown in Fig. 2.4, in which the power factor PF and the form factor ρ are plotted against the normalized inductance value L/L_c . The results are for a constant power load, such as occurs when a pwm regulator follows the filter; Schwarz also gives results for a constant resistance load, which are little different.

The most significant feature of Fig. 2.4 is that, as expected, the PF never achieves the ideal value of unity, but approaches a maximum value of 0.9 as $L \rightarrow \infty$. In fact, this limit is essentially reached for

$L/L_r > 5$. It is noteworthy that the power factor for large L is less than unity in spite of the fact that the input voltage and current are always in the same direction, that is, power is only delivered to the input and is not withdrawn in any part of the cycle. Under these conditions the power factor is less than unity because the form factor is less than $\sqrt{2} = 1.41$, the ideal value for sinusoidal current. In other words, there are harmonics of the line frequency present in the input current waveform which do not contribute to load power.

Although the ideal result cannot be obtained with the rectifier/LC filter circuit, at least Fig. 2.4 permits a suitable compromise choice for the inductance value to be made, as long as a power factor less than 0.9 is acceptable.

2.2 Resistor Emulation

In the rectifier/filter method described in the previous section, the input current to the filter is shaped by an inductor to be approximately sinusoidal over each half cycle of the corresponding sinusoidal voltage, regardless of the output current being drawn from the filter. This circuit is incapable of providing the ideal current waveform.

Since this circuit is usually followed by a dc regulator, improved results at little extra cost can be obtained by augmenting the regulator in such a way that it accepts the rectified sine wave voltage input from the rectifier and at the same time shapes its input current to be the desired half sine waves. Thus, the LC filter function is essentially absorbed into the regulator, as shown in Fig 2.5.

Since the regulator now has to perform two functions, output dc voltage regulation and input current shaping, two control parameters are

needed. These can be achieved by a two-feedback-loop configuration which is a modification of a two-loop circuit that is already gaining favor in dc regulators even when input current shaping is not required. This circuit is often referred to as a current-mode controlled regulator [8,9].

As modified for input current shaping, the block diagram of the regulator is shown in Fig. 2.6. The "major" dc voltage regulation loop is explicit; the "minor" current feedback loop is implicit inside the current generator block. The modification consists merely in the addition of the multiplier block. In its absence, the dc error voltage v_e develops a proportional current in the current generator, which supplies the load and is drawn from the input. Thus, the input current is dc. However, in the presence of the multiplier, the rectified but unfiltered input voltage $e_{r,c}$ is multiplied by the dc error voltage v_e , so that the current $i_{r,c}$ drawn from the input is of the same shape as, and proportional to, the input voltage, which is the desired result. The proportionality factor, the dc error voltage v_e , is automatically adjusted by the voltage feedback loop so that the average value of the $i_{r,c}$ rectified sine wave is the current required by the load at the regulated voltage.

Since the input current waveshape is a scaled version of the input voltage, the input to the regulator appears resistive, which leads to the appellation "resistor simulation" circuit. In principle, it can achieve ideal results, namely, unity power factor and $\sqrt{2}$ form factor; however, practical limitations preclude ideal results.

A hardware realization of the resistor emulation method has been described by Keller and Baker [10], in which a boost converter with

current-mode programming is used to implement the current generator block. Practical limitations include the difficulty of making the converter dynamic range swing to zero input voltage, as required at the beginning and end of each half cycle of input voltage. Also, the bandwidth of the major voltage loop has to be low compared with the line frequency. However, for a line frequency of 20 kHz, this is not a severe limitation.

2.2.1 Three Phase Extension

The simple circuit of the resistor emulation method can be extended to three phases as shown in Fig. 2.7. Three-phase operation has the advantage of zero power ripple delivered to the storage capacitors, and thus zero output voltage ripple. This circuit requires two more slave circuits to be synchronized and running at the same multiplier duty cycle as the master circuit. The slaves must have isolated outputs.

2.3 Pulse-Width-Modulation

This method is attractive because the power factor can be made to be unity regardless of the output voltage; also the lower harmonics of the line current can be reduced by selecting both the number of pulses in a half cycle of the input voltage and the appropriate relation between the pulse widths.

Several techniques are used in pulse-width-modulation (pwm), including sawtooth modulation [11], sub-harmonic overmodulation [12], and firing angle modulation [13]. We have chosen only the most promising of the new pulse-width controlled ac-to-dc converters for comparison with other methods. In other designs, sub-harmonic

overmodulation [12] suffers the introduction of low-frequency harmonics (3rd and 5th); and firing angle modulation [13] requires an extra rectifier, even though its improvement is comparable to sawtooth modulation.

Sawtooth modulation [11] is new in the sense it eliminates the use of auxiliary thyristors and their commutation circuits. A single-phase PWM-controlled rectifier is employed as shown in Fig. 2.8. The operation is as follows: assume at a certain instant, Th_1 is on and the commutating capacitor C has A positive and B negative. When thyristor Th_2 is turned on, the capacitor A positive will reverse bias the thyristor Th_1 and turn the latter off.

The capacitor C discharges through L_1 , D_{L1} and the load and reverses the polarity such that B clamps the supply voltage through D_1 . Then D_2 will begin to conduct also. L_s is included to boost the charge in capacitor C in case the output voltage is zero, in order to ensure successful commutation in the next cycle.

Figure 2.9 shows the sawtooth modulation technique used. The gate signals for the thyristors are obtained by comparing a triangular voltage e_t with a rectified sinusoidal voltage e_s synchronized with the phase of the source voltage. Various output voltages e_d can be obtained by varying the amplitude of e_s or the ratio $e_{s\max}/e_{t\max}$.

This circuit can easily be extended to three-phase operation. This eliminates the low-frequency harmonics, and the remaining (switching frequency) 11th and 13th harmonics can easily be suppressed.

However, the control of each half cycle of the line frequency must be constant in order to avoid the generation of line frequency harmonics. This degrades the response to load and line transients

because of the required low bandwidth of the control loop. Thus a two-stage switch-mode converter is usually required, one to shape the line input current and the other to widen the bandwidth of the line and load regulation from the low-bandwidth dc output of the first converter. Therefore, this method is less cost effective than the two more promising methods for complete solutions that are discussed in Sections 2.1 and 2.2.

2.4 Active Filter

Switched-mode power converters with high switching frequencies and high-gain voltage feedback controls have been used as active filters, in order to reduce low frequency ripple [14,15]. They can be connected either in series or in parallel with the controlled rectifier under study to form one power processing unit to make up an ac-to-dc converter. A more elegant solution is to connect a supercapacitor filter in parallel, as described below.

The circuit diagram is as shown in Fig. 2.10. The dc-to-ac inverter operates at 20 kHz in the reverse direction (by appropriate choice of the switch quadrants) so that the low-frequency ripple of the line input current is zero. The rectified voltage u_R is sensed and a low-frequency current \tilde{I}_c is generated to cause the input current low-frequency ripple to approach zero (through pwm to the switches S_1, S_2, S_3, S_4) while the 20 kHz component is filtered. The voltage across C is $\tilde{u}_C = (1 + F)\tilde{u}_R$, where F is the open-loop gain of the control system. The capacitor C sinks the current \tilde{I}_c which is $(1 + F)$ times larger than \tilde{I}_R . Thus, only a small value of actual C is required.

This scheme can be employed with a three-phase supply if a three-phase rectifier is used. However, the ac port impedance in this case is

very low ($\approx 15 \text{ m}\Omega$). The three-phase design is more difficult to implement and the inverter has to force the current waveshaping. Secondly, an extra inverter is used, which makes this solution less cost-effective.

2.5 Topology Modification

Many good ideas have been presented over the last several years in simplifying switching circuits and improving the general performance of switched-mode power converters. These help to cut the component cost and are particularly useful in very low-power conditioning systems. Some novel topological modifications are discussed as follows.

2.5.1 Direct Source Harmonic-free Rectifier

Figure 2.11 shows the schematic and the components of the power circuit of a 6 kW harmonic-free rectifier using a dc source.

The system is fed with a single-phase ac source V_{ac} . The rectified sinusoidal current waveform is shaped by the dc-to-dc converter consisting of the high-frequency (20 kHz) controllable switch GTO and diode D_1 . On the bridge side of the converter, L_0 and C_0 filter the high-frequency current. The resistor R_0 is used to damp oscillations due to disturbances in the filter when there are sudden changes in power level. On the load side of the converter, L_c and C_c form the "resonant filter." The component values required are much smaller than when only the inductor L_c is used alone. Since the combined impedance of the two elements at the switch frequency is essentially a short circuit, an additional high-frequency filter is therefore needed. It is made up of L_f and C_f . L_f is placed in series with C_c in order to ensure (1) the lowest energy storage requirements in

L_f , and (2) the lowest RMS current through L_f . L_f also provides inductance, which combines with C_o to provide the desired resonance. Thus (1) and (2) are ensured, since the maximum current flowing through L_f is only the peak of the current in the resonant circuit.

In this system the use of the resonant filter modification gives savings in cost, size, weight and also the capability to interface the quick changes in the level of power at the ac port. The results show that when the rectifier is operating at full power, the system has a total harmonic distortion of less than 3% and a power factor very close to 1.0. As a whole, this modification provides many good ideas, especially the resonant circuit to be added to the proposed inductor-input filter solution.

2.5.2 Modified Gating Single-Phase Rectifier Bridge [17]

This configuration is particularly useful in a single-phase thyristor bridge rectifier for power factor improvement and line current harmonic reduction. The thyristor controlled bridge is connected to an inductive load such as a dc motor, as shown in Fig. 2.12. The gating of the thyristors usually provides some control of output voltage e_d . In this modification, it also helps to eliminate the free-wheeling diode required across an inductive load. For instance, assume that at this moment Th_1 and Th_4 are conducting and when the supply voltage V_{ac} crosses zero, the voltage across Th_3 is reversed, biasing Th_4 off. Thus the leg formed by Th_1 and Th_3 forms the free-wheeling path for the load, making the free-wheeling diode across the load unnecessary. Furthermore, by this choice of gating technique, the power factor becomes $PF = \sqrt{2} (1 + \cos \alpha) / [\pi(\pi - \alpha)]^{1/2}$ instead of the conventional value approximately equal to $\cos \alpha$, where α is the firing angle. Thus,

from the equation given above, the power factor can be improved by a suitable choice of firing angle. Also [18] this choice provides a reduction in line current harmonics.

However, despite the above advantages, the topology is only of limited use (limited to single-phase thyristor bridges) and is thus not to be considered as a general solution for different applications.

2.5.3 Modified Thyristor Bridge Converter [18]

A modified three-phase commutated converter for power factor improvement in high power applications is shown in Fig. 2.13. Assume that at this moment current is being conducted through the path formed by $U - Th_1 - Load - Th_6 - V - N$. Auxiliary thyristor Th_8 is fired at angle δ , providing a path from Q to neutral N and turning off thyristor Th_6 ; i.e. the conducting path becomes $U - Th_1 - Load - Th_8 - N$. To see how this can achieve power factor improvement, the rectified waveform is sketched out in Fig. 2.14. By firing Th_8 , the phase-to-phase voltage is diverted to phase-to-neutral voltage thereby maintaining the current positive for only positive voltage (if the voltage and current are not in phase, the power factor will be poor). Thus the auxiliary thyristors improve the power factor by reducing the conduction period for each phase and by transferring to the neutral conductor the ac current during the period when the corresponding ac voltage is negative.

However, by reducing the conduction period of the ac line currents, the modified converter injects a 3rd harmonic into the three-phase system [18]. Also, there is an angle restriction, as can be seen from Fig. 2.12. The auxiliary thyristors will conduct only if the converter firing angle α on the line voltage is smaller than $5\pi/6$ and larger than $\pi/6$. The use of two extra thyristors would mean more firing

circuits, more commutation circuits and hence more cost and thus the modification cannot be a practical solution to our problem.

2.6 High-Frequency Switching

An ac-to-dc converter using high-frequency synthesis can be made to have sinusoidal line input current and unity power factor. This new frequency converter [19] is basically a polyphase input and output converter. A three-phase input and monophasic output converter is shown in Fig. 2.15. The switches are operated at a high frequency, (compared to the power line frequency) sequentially and cyclically. During the k th switching sequences as shown in Fig. 2.16, let the times along which the switches S_1 , S_2 and S_3 are closed be t_1^k , t_2^k , t_3^k where $t_1^k + t_2^k + t_3^k = T_{s\cdot q} = 1/f_{s\cdot q}$ where $T_{s\cdot q}$ = sequence switching period, $f_{s\cdot q}$ = sequence switching frequency. During the k th sequence, the average output voltage is approximately given by $V_{0\cdot v}^k = V_1 t_1^k + V_2 t_2^k + V_3 t_3^k$ where V_1 , V_2 , V_3 are input voltages rotating at an angular frequency ω_i ($\omega_i \gg 2\pi f_{s\cdot q}$) during the k th sequence. Indeed, this equation can be interpreted as giving the average output voltage. $V_{0\cdot v}^k$ is the vector combination of the input voltages V_1 , V_2 and V_3 , weighted by the switching times t_1^k , t_2^k , t_3^k , as depicted in Fig. 2.17. The switching times are defined as:

$$t_1^k = \frac{T_{s \cdot q}}{3} [1 + 2 q \cos(K T_{s \cdot q} \omega_m)]$$

$$t_2^k = \frac{T_{s \cdot q}}{3} [1 + 2 q \cos(K T_{s \cdot q} \omega_m - \frac{2}{3} \pi)]$$

$$t_3^k = \frac{T_{s \cdot q}}{3} [1 + 2 q \cos(K T_{s \cdot q} \omega_m - \frac{4}{3} \pi)]$$

where q = constant

ω_m = angular frequency relative to input

Thus output frequency $\omega_0 = \omega_1 + \omega_m$. But in our case, this is an ac-to-dc converter, i.e. $\omega_0 = 0$ so

$$\omega_m = -\omega_1$$

From Fig. 2.17 it is clear that to generate a dc output voltage and a dc output current, the input ac voltage and current are in phase. This can also be reasoned by converter symmetry, the input voltages acting together with the switching contribute an output frequency (in this case dc output and therefore the output frequency = 0). Therefore the output current carries this zero output frequency. Also because of the symmetry, the input current which contributes to this zero output current frequency will be at exactly the frequency of the input voltage. Thus the power factor can be made to be unity.

For a direct-type converter, the choice of high-frequency synthesis ensures a sinusoidal input current. In practice, the switching frequency cannot be infinitely large. Therefore, very small reactive elements are still needed for high-frequency harmonic reduction.

The monophasic converter of Fig. 2.15 can be easily extended to three output phases, as shown in Fig. 2.18. For details and simulation results, please refer to reference [19]. Moreover, this general technique provides converters having many desirable characteristics,

namely (1) only sinusoidal input and output currents are generated (the latter can be made to be dc) (2) bidirectionality (thus allowing voltage step-up or step-down) (3) input and output frequency, magnitude and phase can be independently controlled (4) reactive power can be generated (5) large reactive elements are not required, and (6) power is processed only once.

However, in the case of the three-phase input converter, the control is more complicated and can only be cost-justified for medium- and high-power applications. Also, the proof of independent control of voltage and power factor is incomplete. In addition, in each particular application, the load must be included as a part of the whole system, in order to determine the limiting conditions. The implementation of the switches is difficult and requires a more complicated control system and thus is an unlikely solution.

2.7 Multiconverter

Multiple converters can be used to improve power factor and reduce supply current harmonics. Figure 2.19 shows a typical two-converter system for a three-phase system.

The real power-reactive power (PQ) diagram as shown in Fig. 20 (a) and (b) can be used to illustrate the power factor improvement provided by the two-converter system. For equal real power P taken by both one-converter (2 PU rating; PU = power units) and two-converter (2 PU rating) systems, the two-converter system (only one of the converters need be fully controlled) takes less reactive power and thus improves the power factor; as $\tan \phi = Q/P$, ϕ becomes smaller and the power factor ($\cos \phi$) increases.

As to the line current harmonic reduction, it depends on the different windings of the three-phase transformers. From Fourier series, for

$$i = \frac{2\sqrt{3}}{\pi} I_d (\cos \omega t - \frac{1}{5} \cos 5 \omega t + \frac{1}{7} \cos 7 \omega t - \frac{1}{11} \cos 11 \omega t + \dots)$$

$$i = \frac{2\sqrt{3}}{\pi} I_d (\cos \omega t + \frac{1}{5} \cos 5 \omega t - \frac{1}{7} \cos 7 \omega t - \frac{1}{11} \cos 11 \omega t + \dots)$$

Thus, through the two-converter transformer connection, the fifth and seventh harmonics etc. are cancelled, while only the harmonics corresponding to $n = 12k \pm 1$ will enter the line current where n - harmonic number, and k is an integer. Thus, the low-frequency harmonics are eliminated and the high harmonics can be easily suppressed in the two-converter configuration.

However, the extra converter, transformer and control scheme make this method less cost competitive and therefore can only be considered as a solution in very high-power applications.

3. CONCLUSIONS

Several methods of shaping the input current waveform in ac-to-dc converters have been reviewed.

The simplest method is certainly the LC filter following the rectifier, discussed in Section 2.1. For 60Hz, or even 400Hz input, the inductor is likely to be prohibitively large. The next simplest method is the resistor emulation approach discussed in Section 2.2, in which the inductor size is determined by the converter switching frequency and not by the line input frequency. Other methods require complicated switch drive algorithms in order to construct the input current waveshape.

With respect to a high-frequency line input, on the order of 20 kHz, the simple LC cannot be discarded so peremptorily, since the inductor size is now comparable with that for the resistor emulation method. In fact, since a dc regulator will normally be required after the filter anyway, the total component count is almost the same as for the resistor emulation method, in which the filter is effectively incorporated into the regulator. However, neither method is ideal: as discussed in Section 2.1, the simple LC filter is incapable of giving a power factor greater than 0.9, and the resistor emulation method incurs practical difficulties because of the need for the switch duty ratio to approach zero.

Further Work

Both the simple LC filter method and the resistor emulation method need to be studied further to determine their performance, with respect to power factor and form factor, as a function of different loads. It is likely that the two methods behave differently.

The resistor emulation method has two other aspects that need to be investigated. One is to determine the degree to which performance falls short of ideal because of practical limitations on the duty cycle range. The other is to determine what special problems may arise if the line input frequency is comparable to the converter switching frequency. It is conceivable that there may be an advantage if the two frequencies are harmonically related, or even equal.

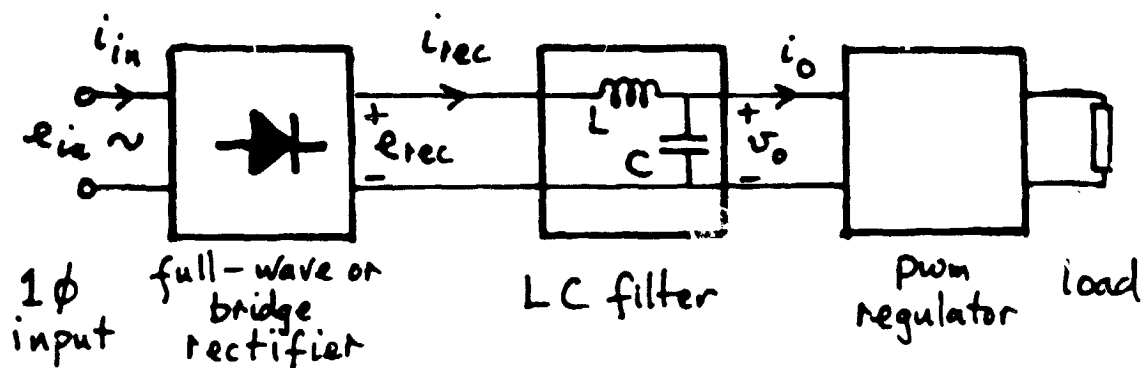


Fig. 2.1 Block diagram, ac-to-dc converter system, LC filter method.

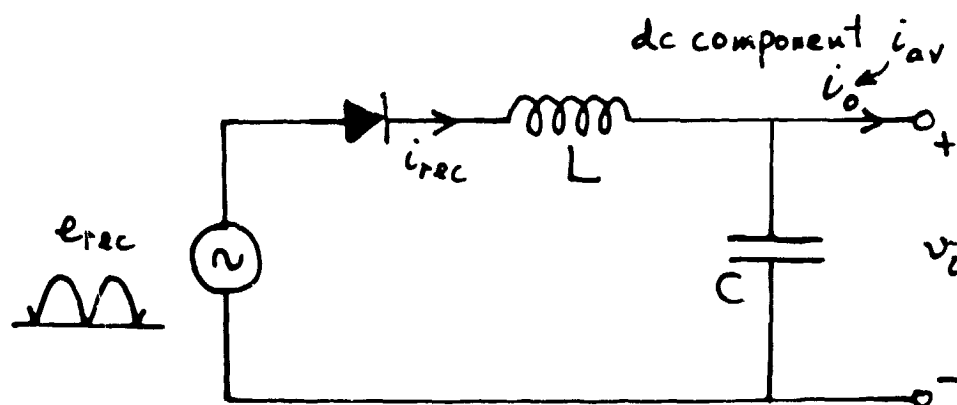


Fig. 2.2 Equivalent circuit, rectifier and filter.

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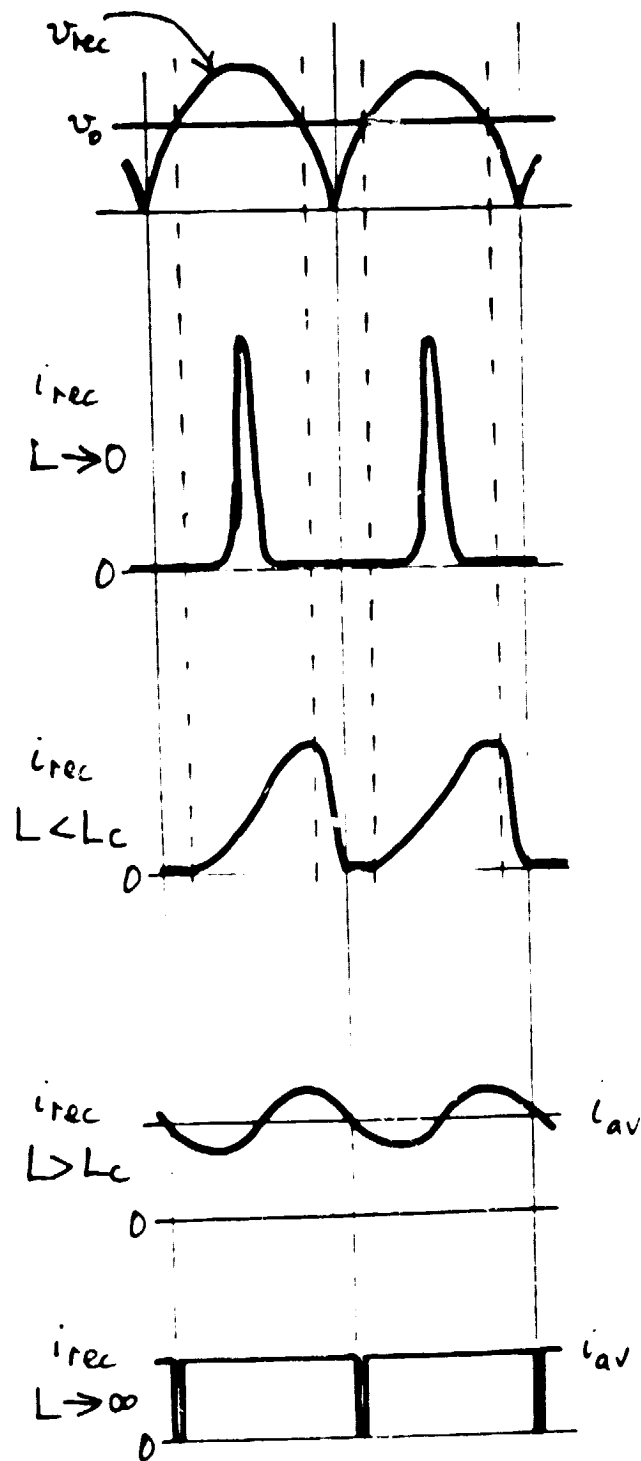


Fig. 2.3 Waveforms, rectifier and filter.

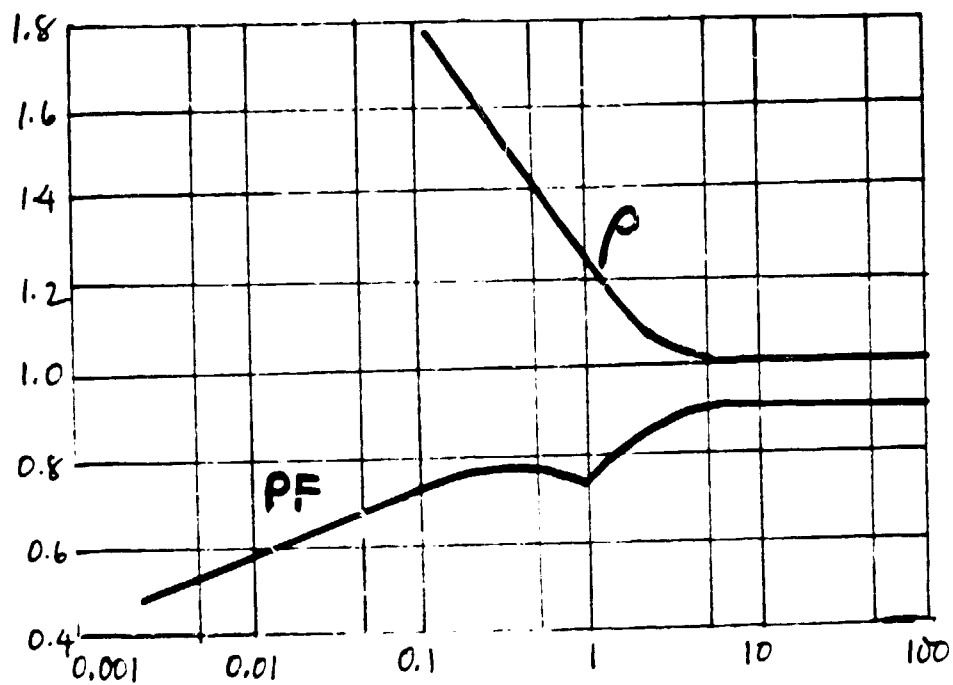


Fig. 2.4 Power factor (PF) and form factor (ρ) as functions of L/L_c (after Schwarz).

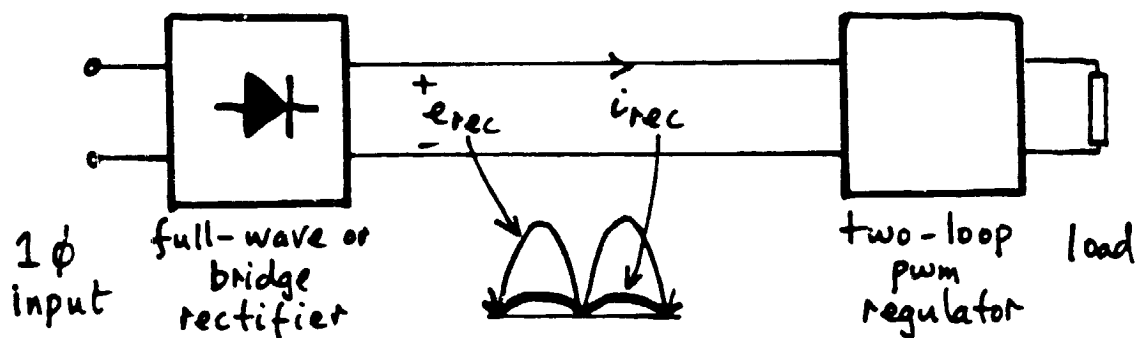


Fig. 2.5 Block diagram, ac-to-dc converter system, resistor emulation method.

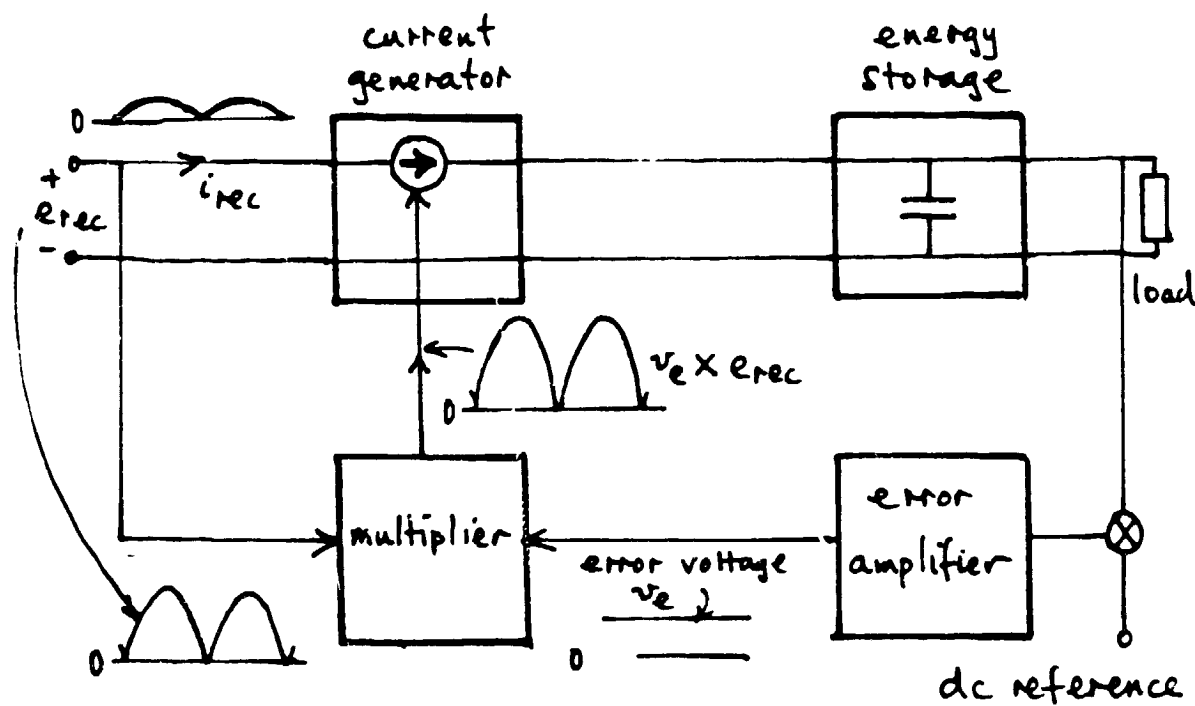


Fig. 2.6 Block diagram, two-loop pwm regulator.

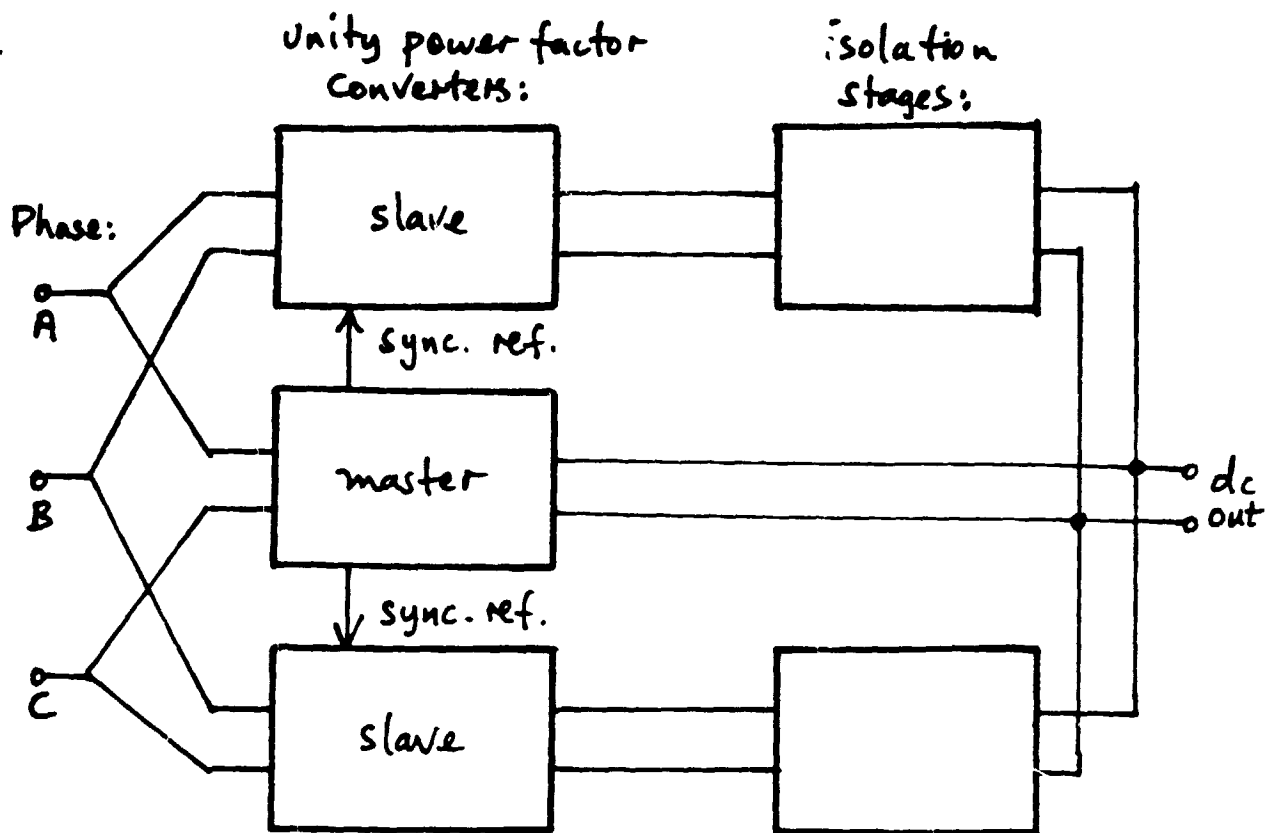


Fig. 2.7 Resistor emulation technique.

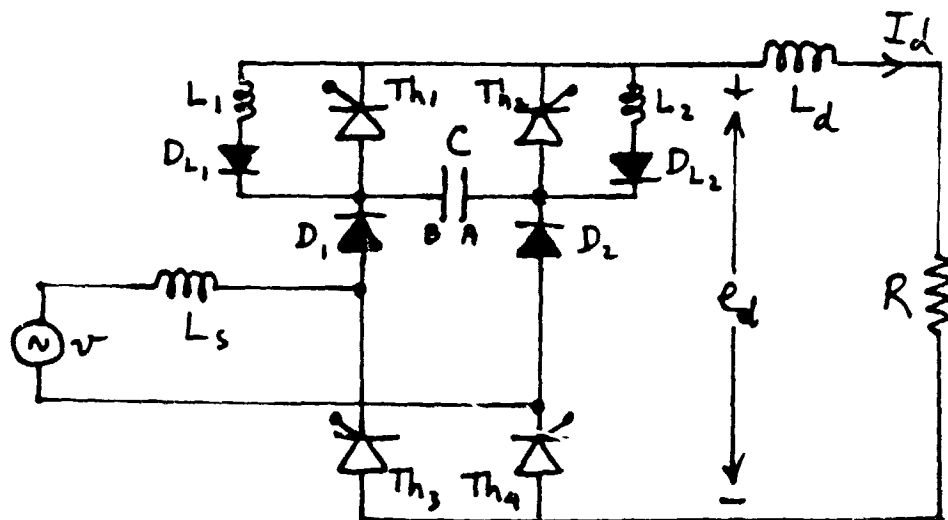
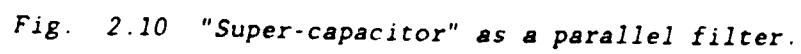


Fig. 2.8 Single-phase pwm-controlled rectifier.



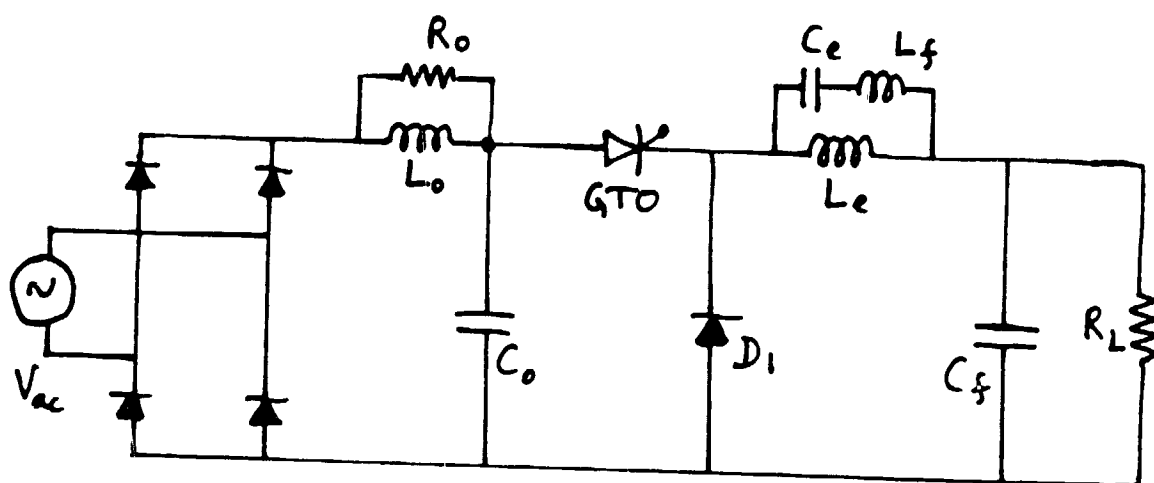


Fig. 2.11 A 6 kW harmonic-free rectifier.

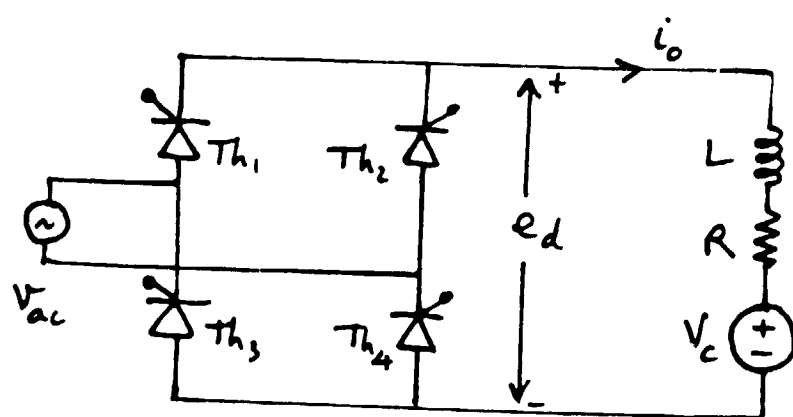


Fig. 2.12 Single-phase thyristor bridge converter.

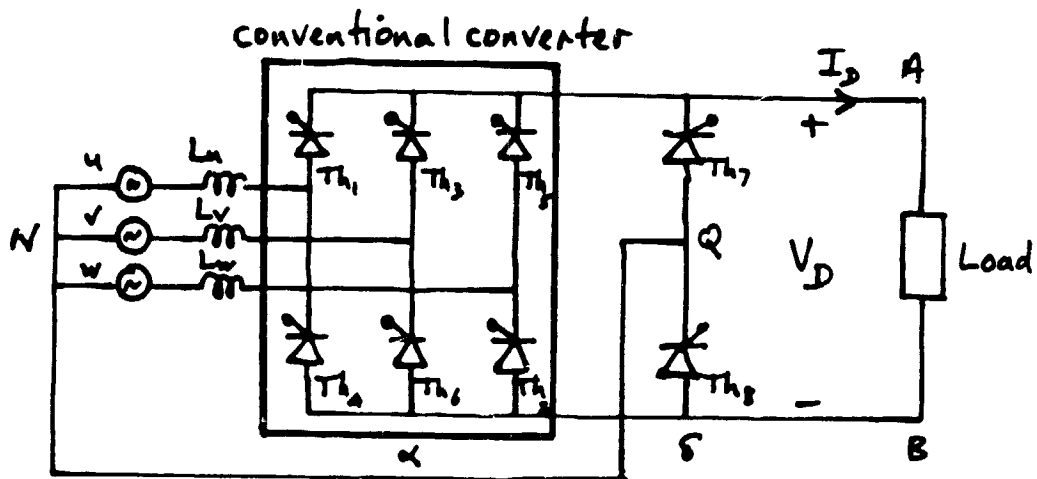


Fig. 2.13 6 pulse bridge, 2 auxiliary thyristor converter.

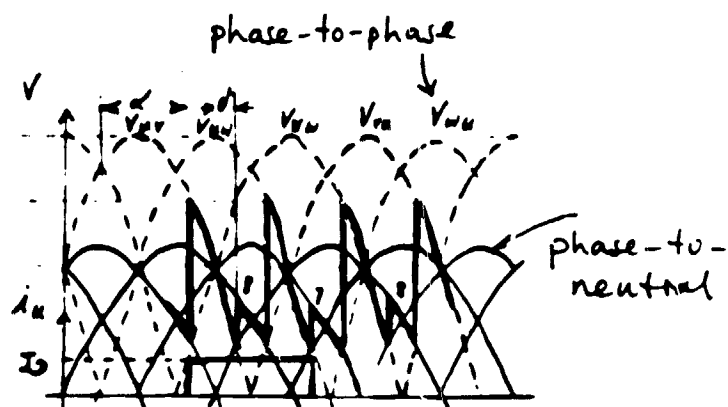


Fig. 2.14 Dc terminal voltage and ac line current during rectifying operation with auxiliary thyristor arbitrary firing angle ω .

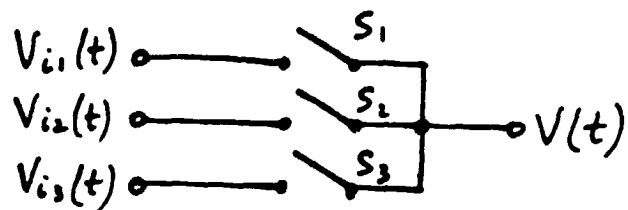


Fig. 2.15 Three-phase, single-phase output converter.

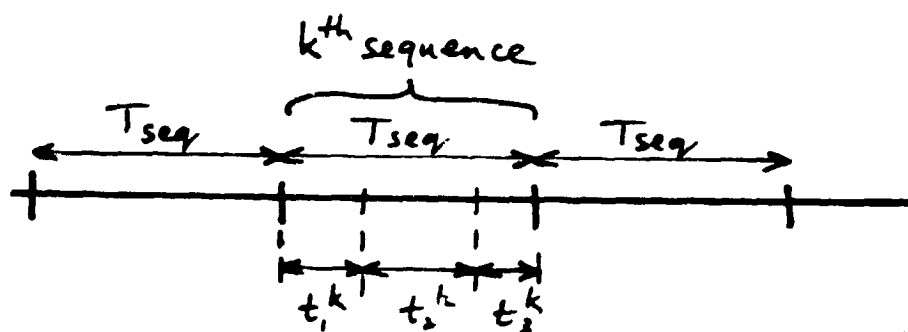


Fig. 2.16 Switching times.

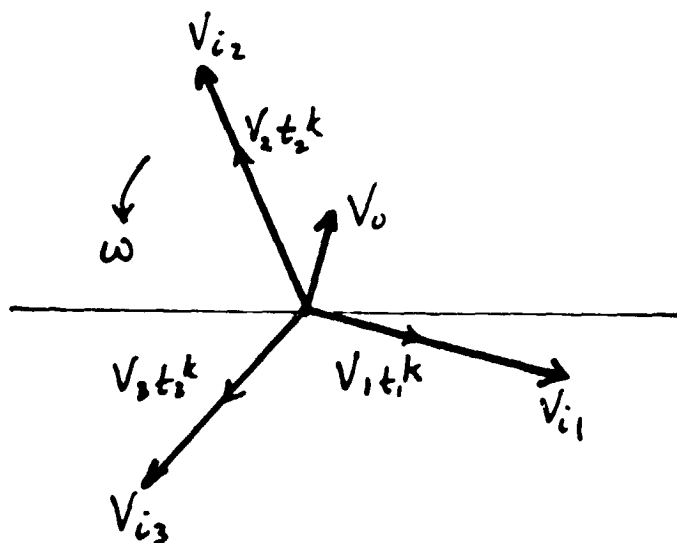


Fig. 2.17 Output voltage synthesis in the three-phase input, single-phase output converter.

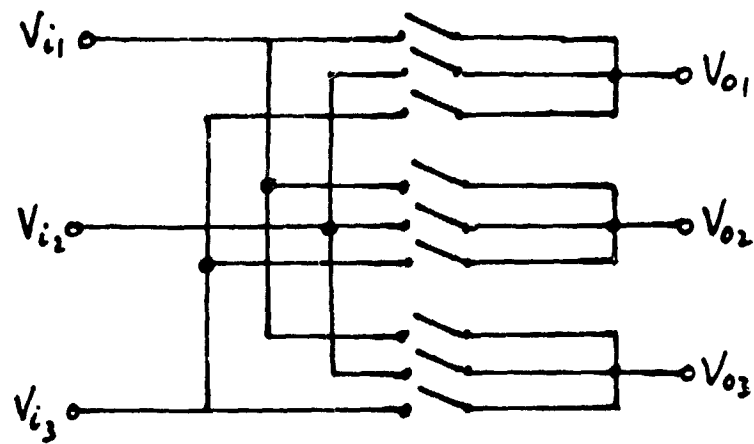


Fig. 2.18 Three-phase to three-phase converter.

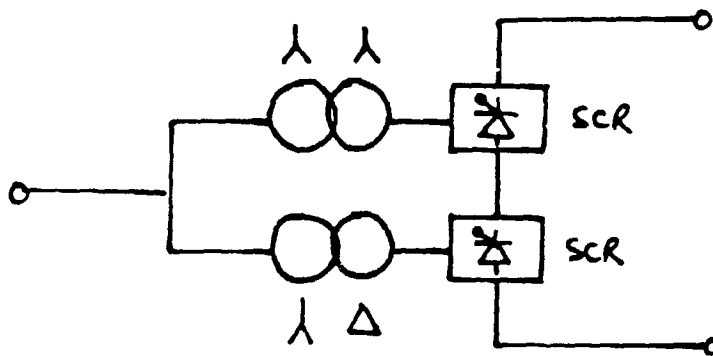


Fig. 2.19 Three-phase two-converter system.

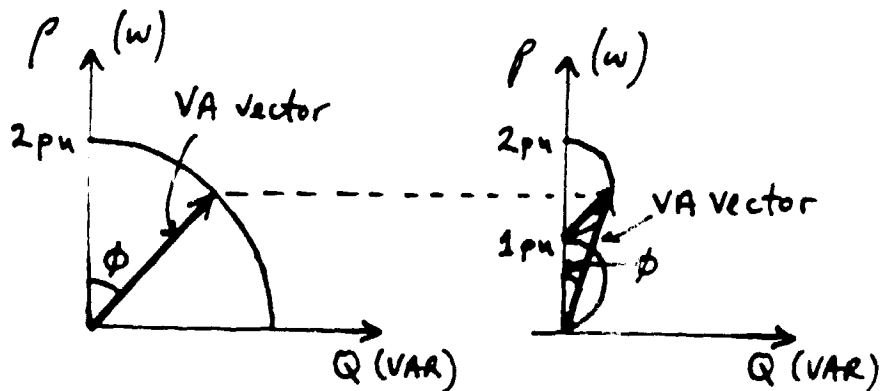


Fig. 2.20 (a) One-converter 2 PU rating. (b) Two-converter 2 PU rating.

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